SYSTEM AND METHOD FOR CANCELING CROSSTALK

CROSS-REFERENCE TO RELATED APPLICATIONS

This application claims the benefit of U.S. Provisional Patent Application Serial

5   No. 60/252,250, filed on November 21, 2000, and entitled "NEXT and FEXT

Cancellation Based on the Common Mode as a Secondary Sensor," which is hereby

incorporated by reference.

BACKGROUND OF THE INVENTION

10   **Field of the Invention**

The present invention generally relates to communication systems, and more

particularly, to a system and method for to canceling crosstalk in a communications

system.

15   **Discussion of the Related Art**

In recent years telecommunication systems have expanded from traditional POTS

communications to include high-speed data communications as well. As is known, POTS

communications includes not only the transmission of voice information, but also PSTN

(public switched telephone network) modem information, control signals, and other

20   information that is transmitted in the POTS bandwidth, which extends from

approximately 300 hertz to approximately 3.4 kilohertz.

New, high-speed data communications provided over digital subscriber lines

(DSL), such as Asymmetric Digital Subscriber Line (ADSL), Rate Adaptive Digital

Subscriber Line (RADSL), High-Speed Digital Subscriber Line (HDSL), *etc.* (more

broadly denoted as xDSL) provide for high speed data transmissions, as is commonly

used in communicating over the Internet. As is known, the bandwidth for xDSL

transmissions is generally defined by a lower cutoff frequency of approximately 30

5   kilohertz, and a higher cutoff frequency which varies depending upon the particular

technology. Since the POTS and xDSL signals are defined by isolated frequency bands,

both signals may be transmitted over the same two-wire loop.

Indeed, twisted pair public telephone lines are increasingly being used to carry

relatively high-speed signals instead of, or in addition to, telephone signals. Examples of

10   such signals are ADSL (asymmetric digital subscriber line), HDSL (High Density

Subscriber Line, T1 (1.544 Mb/s), and ISDN signals. There is a growing demand for

increasing use of telephone lines for high speed remote access to computer networks, and

there have been various proposals to address this demand, including using voice over data

systems to communicate signals via telephone lines at frequencies above the voice-band.

15   The provision in the public telephone network of varied services using such diverse

communications systems imposes a requirement that different and similar systems not

interfere with one another.

In the telecommunication art, the term "crosstalk" refers to interference that enters

a message channel from one or more other channels through a path coupling the message

20   channel with the interfering channels. Crosstalk can create annoyance in a voice system

or errors in a data system. Crosstalk's impact depends upon such factors as the listener's

hearing acuity, the extraneous noise, the frequency response characteristic of the coupling

2

path, and the level of the disturbing signal.

There are generally two types of crosstalk mechanisms that are characterized, one being NEXT and the other being FEXT (far-end crosstalk). NEXT results from a disturbing source connected at one end of the wire pair causing interference in the

5   message channel at the same end as the disturbing source. FEXT is that portion of the disturbing signal propagated to far end of the message channel, that is, the end opposite to the end to which the disturbing source is connected.

Allocations of wire pairs within telephone cables in accordance with service requests have typically resulted in a random distribution of pair utilization with few

10   precise records of actual configurations. In addition, due to the nature of pair twisting in cables, and where cable branching and splicing occurs, a wire pair can be in close proximity to different other pairs over different parts of its length. At a telephone CO (central office), pairs in close proximity may be carrying diverse types of service using various modulation schemes, with considerable differences in signal levels (and receiver

15   sensitivities) especially for pairs of considerably different lengths.

Statistical data has been developed that can be used to estimate crosstalk between services using different pairs of multi-pair telephone cables, for example in terms of BER (bit error rate) based on power spectral density (PSD, for example measured in milliwatts per Hertz expressed in decibels, or dBm/Hz) overlap between the services. However, this

20   statistical data is of limited use in practice in the provision of a new service using equipment connected to a specific wire pair, in view of factors such as those discussed above.

A common feature of the above-mentioned transmission systems is that twisted-pair phone lines are used as at least a part of the transmission medium that connects a central office (e.g., telephone company) to users (e.g., residence or business). It is difficult to avoid twisted-pair wiring from all parts of the interconnecting transmission

5 medium. Even though fiber optics may be available from a central office to the curb near a user's residence, twisted-pair phone lines are used to bring in the signals from the curb into the user's home or business.

The twisted-pair phone lines are grouped in a binder. While the twisted-pair phone lines are within the binder, the binder provides reasonably good protection against

10 external electromagnetic interference. However, within the binder, the twisted-pair phone lines induce electromagnetic interference on each other. This type of electromagnetic interference is generally known as crosstalk interference, which includes NEXT interference and FEXT interference. As the frequency of transmission increases, the crosstalk interference becomes substantial. As a result, the data signals being transmitted

15 over the twisted-pair phone lines at high speeds can be significantly degraded by the crosstalk interference caused by other twisted-pair phone lines in the binder. The problem worsens as the speed of the data transmission increases or as the loop length increases.

As is known, NEXT interference may result from several sources. One such source may be referred to upstream "leak" levels into a DSL downstream service at the

20 customer premise (CP) side. As is also known, there are generally two techniques or approaches to reduce NEXT upstream leak: blind and supervised. Blind source separation is the separation of a composite signal into its constituent component signals,

4

without *a priori* knowledge of those signals. A growing number of researchers have published articles discussing techniques that perform blind source separation (BSS). For example, "Equivariant Adaptive Source Separation", JF CARDOSO & B. LAHELD IEEE Trans. on Signal Proc. , VOL. 44, No. 12, December 1996, and "Adaptive Blind Source Separation for virtually any Probability Density Function", V. ZARZOSO, A. NANDI. IEEE Trans. on Signal processing, Vol. 48, No. 2 February 2000, discuss blind source separation. Both of these articles are hereby incorporated by reference in their entireties. These techniques find use in various applications, including crosstalk removal in multichannel communications, multipath channel identification, and equalization. Many of the BSS techniques require (or assume) a statistical independence between the component signals to accurately separate the signals. Additional theoretical progress in signal modeling has generated new techniques that address the problem of identifying statistically independent signals, which is a recognized problem that lies at the heart of source separation.

In contrast, "supervised" techniques rely on other information sources, collected from physical sensors. For example, "Multichannel Signal Processing for Data Communications in the Presence of Crosstalk", IEEE trans. on Communications, VOL. 38, No. 4, April 1990, and "Frequency selective NEXT Cancellation", J. CIOFFI, ICASSP 2000, VOL. 5, pp. 2841 – 2844, discuss supervised techniques. Both of these articles are hereby incorporated by reference in their entireties.

Accordingly, there is a need to provide an improved system and method for canceling crosstalk interference.

SUMMARY OF THE INVENTION

Certain objects, advantages and novel features of the invention will be set forth in part in the description that follows and in part will become apparent to those skilled in the art upon examination of the following or may be learned with the practice of the invention. The objects and advantages of the invention may be realized and obtained by means of the instrumentalities and combinations particularly pointed out in the appended claims.

To achieve the advantages and novel features, the present invention is generally directed to a method and apparatus for reducing crosstalk in a telecommunication system. Broadly, the present invention utilizes a circuit that obtains a common mode signal from a differential signal of a local loop, to obtain additional information that can be used to better approximate the transmitted signal (by approximating and canceling crosstalk or otherwise). In contrast to some known prior art systems, the system and method of the present invention has no knowledge, nor does it make any assumptions, about the crosstalk or signals from disturber lines.

In accordance with one embodiment of the invention, a modem is provided having improved crosstalk cancellation circuitry for canceling crosstalk received on a local loop (or otherwise estimating the remotely transmitted signal) carrying modem communications. The modem includes a first input for receiving a signal carried on the loop and a second input for receiving a signal obtained from the common mode of the local loop. The modem further includes processing circuitry configured to either reduce

6

crosstalk present in the signal carried on the local loop, or to otherwise closely approximate the remotely transmitted signal. The modem accomplishes this through computations based upon a first signal carried obtained from the local loop and a second signal obtained from the common mode of the local loop. Using these signals allows the

5   processing circuitry of the modem to obtain more accurate results. In this regard, mathematical equations or expressions may be generated that model the received signal in terms of the remotely transmitted signal, the impulse of the local loop, the signal on one or more disturber lines, the impulse response of the crosstalk coupling, white noise, etc. The modem learning phase may be utilized to compute some of these terms, using blind,

10   semi-blind, or supervised techniques. The use of a signal obtained from the common mode can provide an additional equation, which can be used and processed to provide more accurate computation results. Ultimately, this provides better crosstalk cancellation or otherwise better approximation of the remotely transmitted signal.

       In accordance with another embodiment of the present invention, a method for

15   reducing crosstalk on the local loop is also provided. The method obtains a first signal from the local loop and obtains a second signal from the common mode. Then, the method uses the first and second signals to compute a measure of crosstalk present on the local loop, and subtracts the measure of crosstalk from the first signal.

       In an alternative embodiment, a similar method is provided. However, instead of

20   computing a measure of crosstalk and subtracting the measure from the first signal (i.e., the received signal), the alternative embodiment approximates the transmitted signal. One way this approximation may be carried out is to utilize a covariance matrix (see

7

Equation 28 below) and statistical probabilities to estimate or approximate the transmitted signal.

DESCRIPTION OF THE DRAWINGS

5   The accompanying drawings incorporated in and forming a part of the specification, illustrate several aspects of the present invention, and together with the description serve to explain the principles of the invention. In the drawings:

FIG. 1 is a diagram illustrating a basic DSL environment, including a customer premises and central office as is known in the prior art;

10   FIGS. 2A and 2B illustrate the manner in which multiple local loops are bundled, which may result in cross-coupling and crosstalk;

FIGS. 3A and 3B are diagrams illustrating various sources of crosstalk and NEXT disturbance in a bundled line;

FIG. 4 is a diagram similar to FIG. 3A, but illustrating the use of a common mode

15   signal in accordance with the invention;

FIG. 5 is a block diagram illustrating certain components in the receiver portion of a DSL modem constructed in accordance with one embodiment of the invention;

FIG. 6 is a block diagram illustrating certain components in the receiver portion of a DSL modem constructed in accordance with another embodiment of the invention;

20   FIG. 7 is a flowchart illustrating the top level operation of one embodiment of the invention;

FIG. 8 is a flowchart illustrating the top level operation of an alternative embodiment of the invention; and

FIG. 9 is a circuit diagram illustrating one circuit configuration that may be used to obtain a common mode signal.

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DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Having summarized various aspects of the present invention, reference will now be made in detail to the description of the invention as illustrated in the drawings. While the invention will be described in connection with these drawings, there is no intent to

10　limit it to the embodiment or embodiments disclosed therein. On the contrary, the intent is to cover all alternatives, modifications and equivalents included within the spirit and scope of the invention as defined by the appended claims.

As will be appreciated by persons skilled in the art, the concepts and teachings of the present invention may be applied to many types of telecommunication systems.

15　However, the preferred embodiment and environment of the present invention applies to digital subscriber line (DSL) systems. As is known, the length of the local loop is a significant and limiting factor for DSL systems. Providing an improved system and method for reducing crosstalk can operate to extend the length or reach of the local loop in a DSL system. Although persons skilled in the art will understand the components and

20　operation of a DSL system, reference is briefly made to FIG. 1, which summarizes the basic DSL environment.

9

In this regard, and as is known, the POTS frequency band generally extends from approximately DC to approximately 3.4 kilohertz. The DSL frequency band generally extends from approximately 30 kilohertz up to an upper boundary, which varies depending upon the particular digital subscriber line technology utilized. A separation or guard band provides some margin of signal separation, and thus noise and cross-talk immunity between signal transmissions between the POTS and DSL frequency bands.

Referring now to FIG. 1, a prior art communication system is shown. Specifically, FIG. 1 illustrates communication between a central office 20 and a customer premises 22 by way of local loop 24. At the central office 20, additional circuitry is provided. Generally, a line card 40 (denoted by dashed lines) containing line interface circuitry is provided for electrical connection to the local loop 24. One or more xDSL modems, including both receiver and transmitter circuitry, are embodied in a DSL access multiplexer (DSLAM), denoted by block 42. This circuitry is directly connected for communications across the local loop 24. The remaining POTS circuitry may be separated, or buffered, from the local loop 24 by a POTS filter contained within a POTS splitter 30. As is also known, the POTS splitters 30 and 44 may or may not be required, depending upon the transmission protocol. For example, G.Lite and ADSL.Lite do not require POTS splitters and filters. For this reason, these components have been illustrated in dashed line.

While the customer premises 22 may be a single dwelling residence, a small business, or other entity, it is generally characterized as having POTS equipment, such as a telephone 26, facsimile machine 29, PSTN modem 27, *etc.* The customer premise 22

may also include an xDSL communication device, such as an xDSL modem 28. When an

xDSL service is provided, a POTS splitter 30 may be interposed between the POTS

equipment (*e.g.*, 26, 27, 29) and the local loop 24. As is known, the POTS splitter 30

includes a low-pass filter having a cut-off frequency of approximately 4 to 10 kilohertz,

5   in order to filter high frequency transmissions from the xDSL communication device 28

and protect the POTS equipment.

As illustrated, and as is known, in a typical customer premises, a plurality of RJ11

jacks 47, 48, and 49 are provided throughout the customer premises. These are typically

interconnected, and originate from the point of entry into the customer premises. In the

10   illustration of FIG. 1, each may originate from a POTS filter (within POTS splitter 30),

which may be provided at the point of entry. Any POTS communication device may be

directly plugged into any of these predisposed jacks 47, 48, and 49.

DSL communications typically use two pieces of equipment, one on the customer

end, and one at the Internet Service Provider, telephone company or other provider of

15   DSL services. At the customer's location there is a DSL transceiver (sometimes referred

to as a DSL modem and other times referred to as an ATU-R), which may also provide

other services. The DSL service provider has a DSLAM 42 to receive customer

connections.

The DSL modem 28 is the point where data from the user's computer or network

20   is connected to the DSL line (i.e., the local loop 24). The transceiver 28 can connect to a

customer's equipment in several ways, though most residential installation uses USB or

10-baseT Ethernet connections. While most of the DSL transceivers sold by ISPs and

telephone companies are simple transceivers, the devices used by businesses may combine network routers, network switches or other networking equipment in the same platform.

The DSLAM 42 at the access provider is the equipment that really allows DSL to happen. A DSLAM 42 takes connections from many customers and aggregates them onto a single, high-capacity connection to the Internet. DSLAMs are generally flexible and able to support multiple types of DSL in a single central office and different varieties of protocol and modulation -- e.g., both CAP (Carrierless Amplitude Phase) and DMT (Discrete Multi-Tone), for example -- in the same type of DSL. In addition, the DSLAM 42 may provide additional functions including routing or dynamic IP (Internet Protocol) address assignment for the customers.

The DSLAM 42 provides one of the main differences between user service through DSL and through cable modems. Because cable modem users generally share a network loop that runs through a neighborhood, additional users mean lowered performance in many instances. DSL provides a dedicated connection from each user back to the DSLAM 42, meaning that users won't see a performance decrease as new users are added--until the total number of users begin to saturate the single, high-speed connection to the Internet. At that point, an upgrade by the service provider can provide additional performance for all the users connected to the DSLAM 42.

As shown in FIG. 2A, multiple loops from various customer premises 102, 104, 106, 108, 110, are typically grouped in a common bundle 120 (sometimes called a binder), through which they are routed to a central office 114. As shown in FIG. 2B, the

various lines are capacitively-coupled, due to their close proximity within the bundle 120. This coupling permits crosstalk. As is known, crosstalk is a limiting factor in the wideband DSL standards that are currently evolving. Due to the capacitive coupling between adjacent wires in a bundle, data transmissions on one line 130 may interfere with

5   data transmissions on another line.

Reference is now made to FIGs. 3A and 3B, which more particularly illustrate the various sources of crosstalk in a telecommunications system. For example, FIG. 3A illustrates only two of the lines 130 and 132, which may be present within a bundle 120. As previously mentioned, the crosstalk noise may be separated into two identifiable

10   components, NEXT and FEXT. Assume that a DSL modem at CP1 102 as a disturber source that interfers or injects crosstalk onto the loop for CP2 104. The NEXT component 140 of this crosstalk is represented by the symbol $g(t)$ (the impulse response of the NEXT cross-coupling between the disturber line and the disturbed line). Likewise, the FEXT component 142 is represented by the symbol $l(t)$ (the impulse response of the

15   FEXT cross-coupling between the disturber line and the disturbed line). In addition, the loop itself may be characterized by an impulse response, represented by the symbol h. As is further known, the NEXT component is typically a much more significant crosstalk component (especially for long loops), since the NEXT disturber source is typically much closer to the disturbed receiver.

20   Crosstalk noise may be reduced by subtracting the induced crosstalk. Since it is impossible to determine an exact quantization of crosstalk that is induced, an estimation or approximation of the crosstalk may be subtracted from the signal at a disturbed

13

receiver. Statistical properties of the crosstalk might be also used to increase the accuracy of a straight main signal estimation. As will be appreciated, better noise immunity is obtained, as crosstalk is more accurately measured or estimated. Consequently, the reach of the local loop may be extended for DSL communications.

5   Ignoring FEXT, a signal received at a receiver may be expressed as:

$$r[n] = (h \otimes e)[n] + (g \otimes u)[n] + w[n], \qquad \text{Equation 1}$$

where, r denotes the received signal, h denotes the impulse response of the loop, e denotes the transmitted signal that is desired to be recovered, g(t) denotes a impulse response of the cross-coupling component between the disturber line and the disturbed

10   line, $u[n]$ denotes the signal on the disturber line, $w[n]$ denotes white noise, n represents discrete time samples, and the $\otimes$ symbol represents a convolution operation.

Reference is now made to FIG. 4 which is a diagram similar to the diagram of FIG. 3A, but illustrating one embodiment of a system constructed in accordance with the invention. In this regard, the present invention is directed to a system and method for

15   canceling crosstalk (or otherwise reducing the effects of crosstalk) on a local loop 130, which carries a first signal $r(t)$. A fundamental feature of the present invention is the utilization of a common mode signal, obtained from the local loop, to generate a second signal $p(t)$. Both of these signals ($r(t)$ and $p(t)$) are utilized to create multiple independent equations that can be solved to either directly approximate the remotely transmitted signal

20    $e(t)$, or to approximate a measure of crosstalk on the local loop, and from the measure of crosstalk approximate the transmitted signal $e(t)$.

As summarized above, two modes have access to the victim modem 200. One is the classical differential mode that brings the DSL service and the other one is the common mode. It is assumed that both modes experience NEXT due to at least one disturbance source. For the moment, we consider only one disturber $u(t)$ (the extension to

5 two or more disturbers is straightforward). NEXT into the common mode is due to the coupling between the disturber pair (differential pair) and the common mode. NEXT into the victim modem 200 is due to the coupling between the disturber line 132 and the mode supporting DSL service, *i.e.* the differential mode. The DSL downstream signal $e(t)$ experiences propagation along the loop. The same DSL downstream signal is coupled

10 into the common mode. The victim modem 200 receives two signals $r(t)$ and $p(t)$ (or $r[n]$ and $p[n]$, after sampling), respectively on the differential and on the common modes. These signals share filtered versions of the DSL downstream signal $e[n]$ and the disturber signal $u[n]$.

The present invention uses both signals $r[n]$ and $p[n]$ to develop an accurate

15 estimate of the downstream DSL signal, $\tilde{e}[n]$.

In FIG. 3A, a cross-coupling components $g(t)$ (for NEXT) and $l(t)$ (for FEXT) were discussed. Similar cross-coupling components are present in the system of FIG 4. If f, g, h, and l designate impulse responses of propagation and coupling effects as illustrated in FIG. 4. If we neglect the FEXT from 132 into 130, both received signals

20 take the following form:

$$p[n] = (f * u)[n] + (k * e)[n] + w[n] \qquad \text{Equation 2}$$

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where k[n] is the coupling between the emitted differential mode of 130 into the received common mode of 130. Indeed, $(k*e)[n]$ represents the leak of the emitted signal into the common mode.

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$$r[n]=(h*e)[n]+(g*u)[n]+v[n] \qquad \text{Equation 3}$$

w[n] and *v[n]* are two independent processes. *w[n]* is as usually assumed (into the differential mode) a white gaussian noise. The signal *w[n]* exhibits a white noise component *η[n]* plus RFI perturbations *rf[n]* due to the "antenna" behavior of the

10   common mode:

$$w[n] = \eta[n] + rf[n] \qquad \text{Equation 4}$$

15   Consistent with the scope and spirit of the present invention, there are a variety of approaches to "solving" the equations presented above. One way would be to compute a measure of the crosstalk present in the received signal *r(t)*, and then subtract that measure from the received signal to approximate the transmitted signal *e(t)*. Another approach would be to use a covariance matrix and statistical probabilities to directly approximate

20   the remotely transmitted signal *e(t)*. In these approaches, the modem learning phase may be utilized to obtain certain information about the variables of Equations 2 and 3. Also, blind, semi-blind, and/or supervised techniques may also be utilized. Such techniques are known and therefore need not be described herein.

The problem is based on Equations 2 and 3 that may be rewritten in the time domain

25   as a noisy convolutive mixing of 2 independent sources:

$$\binom{r}{p}[n] = \begin{pmatrix} h & g \\ k & f \end{pmatrix} * \binom{e}{u}[n] + \binom{v}{w}[n] \qquad \text{Equation 5}$$

16

In the frequency domain, for bin q and symbol t, we have a linear mixing problem:

$$\binom{R}{P}[q,t] = \begin{pmatrix} H & G \\ K & F \end{pmatrix}[q]\binom{E}{U}[q,t] + \binom{V}{W}[q,t] \qquad \text{Equation 6}$$

5   In order to reduce the contribution of the RFI into $w[n]$ (see Equation 3), a pre-processing

of the common mode $p[n]$ is required, such as a linear filtering, notch filtering of the RFI

and (or) matched filter to the spectral density. We call $p'[n]$ the preprocessed common

mode:

10    $$p'[n] = \aleph(p[n]) \qquad \text{Equation 7}$$

Then we have, still in the frequency domain:

15    $$\binom{R}{P'}[q,t] = \begin{pmatrix} H & G \\ K' & F' \end{pmatrix}[q]\binom{E}{U}[q,t] + \binom{V}{W'}[q,t] \qquad \text{Equation 8}$$

Any semi-blind method based on 2 sensors may be applied either in the time domain or in

the frequency domain.

In keeping with the description of FIG. 4, in a system constructed in accordance

20   with the invention, a site (typically a customer premises) receives a local loop 130. From

the local loop, two signals $r(t)$ and $p(t)$ are obtained. The first signal is the signal $r(t)$

directly received by a DSL modem 200. As is known, this is a differential signal. As is

also known, this loop 130 may also carry a POTs signal, since POTS signals and DSL

signal are carried in separate frequency bands.

25   The second signal is the common mode signal. This signal is obtained circuitry

260, which will be described in more detail in connection with FIG. 9. Significantly, this

circuitry obtains the common mode signal, without deteriorating or otherwise affecting

17

the first signal. As will be discussed in more detail below, the DSL modem 200 includes

circuitry 250 for processing the received and common mode signals $r(t)$ and $p(t)$ to

determine the transmitted signal $e(t)$.

Reference is made to FIG. 5, which is a diagram illustrating certain components of

5   a modem receiver 200, which may be used to implement the crosstalk cancellation of the

present invention, in the time domain. Of course, the modem 200 will include many

other components, which need not be illustrated herein, as they will be understood by

persons skilled in the art. Instead, FIG. 5 illustrates those components, which most

impact the implementation of the present invention. In a first data stream, a received

10   signal ($r(t)$) is sampled and converted by analog to digital (A/D) converter 202 into a

sampled signal ($r(n)$). The signal output from the A/D 202 is then passed through a low-

pass filter 204, which prevents aliasing.

In a second data stream, a second signal ($p(t)$) is sampled and converted by A/D

converter 206 into a sampled signal ($p(n)$). The signal output from the A/D 206 is then

15   passed through a low-pass filter 208, which prevents aliasing of the second signal. Each

data stream is then processed separately by time-domain equalizers 210 and 212.

Thereafter, the respective data streams are processed in the time domain processing

circuitry 214. This circuitry performs the mathematical processing, including the

convolution and de-convolution operations, for solving the equations of Equations 2 and

20   3, to determine the crosstalk components and/or the original transmitted signals. As

described herein, the "solutions" to these equations may involve computing a measure of

the crosstalk on the first local loop 130, and subtracting this component from the received

18

signal *r(t)*, in order to compute (or closely approximate) the remotely transmitted signal *e(t)*. Alternatively, the solution to these equations may be to use a covariance matrix and statistical probabilities to directly estimate the transmitted signal *e(t)* (*i.e.*, without first computing and subtracting a measure of crosstalk).

5   Reference is made to FIG. 6, which a diagram similar to FIG. 5, but illustrates certain components of a receiver 300, which may be used to implement the crosstalk cancellation of the present invention in the frequency domain. In short the A/Ds 302 and 306, the filters 304 and 308, and the time domain equalizers 310 and 312 function in the same fashion as described in connection with FIG. 5. The output of the time domain

10   equalizers (TEQs) 310 and 312, however, are directed to FFTs 322 and 324, respectively, which compute the FFTs of the TEQ outputs. There are many well-known techniques and algorithms for computing the FFT of a digital data stream, so these details need not be described herein. Significantly, the FFTs 322 and 324 convert the time-domain data streams into the frequency domain data stream. These data streams may then be

15   processed by the frequency domain processing circuit 314. As is known, the mathematical intensity of solving the equations of Equations 2 and 3 may be simplified by processing in the frequency domain, as time domain convolution operations become frequency domain multiplication operations instead. Again, the performance of the mathematical operations is within the level of one possessing ordinary skill in the art, so

20   it need not be described herein.

19

It should be appreciated that the functional blocks of the diagrams illustrated in FIGS. 5 and 6 may be implemented in hardware, software, or a combination of the two, including micro-coded hardware.

In view of the foregoing discussion, it will be appreciated that one aspect of the present invention embodies a novel method for reducing crosstalk in a telecommunication system. Reference is now made to FIG. 7, which is a flowchart that illustrates the principal steps of this method. Specifically, during a training or learning phase of a modem, certain variables and/or loop characteristics (e.g., loop impulse response) are computed and/or estimated (step 402). These computed/estimated values are later used during modem operation. During this later modem operation, a first signal $r(t)$ is received on a local loop (step 404). At the same time, a second signal $p(t)$ is obtained from the common mode (step 406). Generating/obtaining the second signal $p(t)$ in this way provides a measure of redundancy, which is used to establish multiple, independent equations in the form of Equations 2 and 3 above, which can be solved to compute (or closely estimate) the crosstalk (step 406). After the crosstalk is computed, is may be subtracted from the received signal to obtain the transmitted signal (step 408).

Reference is made briefly to FIG. 8, which illustrates an alternative embodiment of the invention. As shown, the method illustrated in FIG. 8 is similar to the method of FIG. 7, except that the measure of crosstalk itself is not computed. Nor is a measure of crosstalk subtracted from the received signal to approximate the transmitted signal. Instead, statistical probabilities may be used to estimate the crosstalk covariance matrix, which may be used to estimate the transmitted signal (step 412).

Reference is now made briefly to FIG. 9, which is a circuit diagram of a portion of the DSL modem 200 that contains a circuit that may be utilized to obtain the common mode signal. It should be appreciated that numerous circuits may be employed consistent with the scope and spirit of the present invention. The circuit of FIG. 9, however, is

5   provided herein purely for purposes of illustrating one embodiment. Since the details of the circuit of FIG. 9 are not deemed to be limiting upon the present invention, its structure and operation will not be described in detail herein. In the embodiment illustrated, the values of the various components are preferred to be as follows: T1 is a C1905 with a 474 μHenry winding. T2 is a C2208 with a 2mHenry winding. C1 and C2 are both 68

10   nanoFarad capacitors. C3 is a 680 pico Farad, 250 volt capacitor. C4 is a 1.5 nanoFarad. Finally, R1 and R2 are both 49.9 ohm resistors.

In the configuration of FIG. 9, the circuit obtains the common mode signal p(t) as shown. In addition, the circuit configuration performs a band pass filtering operation, having a pass band extending from 138 kHertz to 1.1 Mhertz. This band pass filter

15   contains a second order high-pass filter with a corner frequency at 138 kHertz and a first order low-pass filter with a corner frequency at 1.1 MHertz. Advantageously, the band pass filter prevents out of band signals from saturating the analog front end (AFE).

As previously mentioned, one approach to "solving" Equations 2 and 3 is to make

20   certain assumptions. For example, it may be assumed that the FEXT component is significantly smaller than the NEXT component. Making this assumption reduces Equation 3 to:

21

$$r[n] = (h \otimes e)[n] + (\xi \otimes p)[n] + \eta[n], \qquad\qquad \text{Equation 9}$$

where $\xi[n]$ denotes an equivalent coupling and $\eta[n]$ denotes a white noise. In addition, by ignoring FEXT, we can make the assumption that $p[n]$ is substantially uncorrelated with $e[n]$. During a silent phase (i.e., $e[n]=0$), $\xi[n]$ may be learned either in the time or frequency domain, using a minimum mean square deconvolution or adaptive filtering. Later, during data mode, the term $(\xi \otimes p)[n]$ may be subtracted from the received signal $r[n]$ in a block or adaptive fashion, eventually including decision feedback. To improve the adaptation of ξ and to reduce the interference from out of band signals, $p[n]$ should be band pass or high pass filtered to the adaptive filter representing ξ.

The equations expressed above may be readily extended to account for a plurality of NEXT and FEXT sources. In the case of a plurality of NEXT sources indexed by $j, 1 \le j \le N$, the received DSL differential mode signal $r[n]$ and common mode signal $p[n]$ take the following form:

$$r[n] = (h*e)[n] + \sum_{j=1}^{N}(g_j * u_j)[n] + v[n] \qquad\qquad \text{Equation 10}$$

$$p[n] = \sum_{j=1}^{N}(f_j * u_j)[n] + (k*e)[n] + w[n] \qquad\qquad \text{Equation 11}$$

Now the coupling impulse responses bear relevant subscripts, Equations 9 and 10. As a matter of fact based on spectral density and probability density equivalence, it is possible to rewrite the above equation, as if only one "global "NEXT source $\tilde{u}[n]$ was present:

$$r[n] = (h*e)[n] + (\tilde{g}*\tilde{u})[n] + v[n] \qquad\qquad \text{Equation 12}$$

$$p[n] = (\tilde{f}*\tilde{u})[n] + (k*e)[n] + w[n] \qquad\qquad \text{Equation 13}$$

22

Now the problem consists in recovering *e[n]* based on the signals: {*r[n], p[n]*}, and constitutive Equations 9 and 10. Again, these equations may be solved using methods that are known or readily understood by persons skilled in the art. In this process, the modem learning phase may be used and/or certain approximations may be made. The following discussion explores some of the mathematical methods and techniques, which may be utilized in this process. The following is presented only for purposes of illustration, and it not considered to be necessary for, or limiting upon, the present invention.

We propose hereafter a model of several different Xtalks, which have an equivalent spectral density and probability density and, moreover is very simple. We use the following notations: $p_q[nT_0], 1 \le q \le Q$ designates Q crosstalks, which are by essence uncorrelated, zero mean, white and not necessarily Gaussian. Their spectral density and variances are respectively denoted by $\dot{\gamma}_q(v)$ and σ_q^2. Finally $\dot{g}_q(v)$ represents both the spectral shaping and the coupling frequency transfer functions. Its inverse Fourier transform is written $g_q[nT_0]$. Let us consider now the global signal due to the contribution of all the Xtalks :

$$y[nT_0] = \sum_{q=1}^{Q} g_q[(n-s)T_0]p_q[sT_0] \qquad \text{Equation 14}$$

Because of the limit central theorem, the global signal is Gaussian. Its spectral density is equal to:

$$\dot{\gamma}_y(v) = \sum_{q=1}^{Q} \sigma_q^2 |g_q(v)|^2 \qquad \text{Equation 15}$$

Equation (14) may be rewritten in the following manner :

23

$$\dot{\gamma}_y(v) = (\sum_{q=1}^{Q}\sigma_q^2)\sum_{q=1}^{Q}\alpha_q|g_q(v)|^2 \qquad \text{Equation 16}$$

If we introduce filters of frequency gain $\dot{g}(v)$ (the solution is not unique) such that :

$$|\dot{g}(v)|^2 = \sum_{q=1}^{Q}\alpha_q|g_q(v)|^2 \qquad \text{Equation 17}$$

and a global variance

$$\sigma^2 = \sum_{q=1}^{Q}\sigma_q^2 \qquad \text{Equation 18}$$

then the spectral density of the global signal is equal to :

$$\dot{\gamma}_y(v) = \sigma^2|\dot{g}(v)|^2 \qquad \text{Equation 19}$$

Therefore, to summarize the foregoing: a summation of white and uncorrelated crosstalks have the same spectral density of a unique white disturber $p[n]$ with a coupling channel frequency transfer function $\dot{g}(v)$ satisfying (15) and with a variance equal to (17). A consequence of the above property is that it is not necessary to identify the various crosstalks (from various sources) one by one to reduce their global effect.

Blind methods may be used either to equalize or to estimate $g^{(i)}$. As is known, blind equalizers make use of linear prediction, CMA, MMA RCA algorithms. Blind identification will be based on subspace methods. With the blind equalizers, compensation of the crosstalks can be made. Based on the estimated channels, global deconvolution can be made.

24

As will be appreciated, a semi-blind method may also be employed to estimate certain variables. As is known, using a semi-blind estimation technique, certain information may be approximated or calculated during the learning phase of the modem. Since various methodologies are known for implementing semi-blind computations and

5 estimations, they need not be described herein.

As discussed above, the computations carried out by the present invention may be performed in either the time domain or the frequency domain. Having discussed a time domain approach above, reference will now be made to a frequency domain approach.

The synthesis hereafter makes an explicit use of the bin rank q and the symbol

10 time index t. All of these methods are based on the "double path" that stems from the oversampling by a factor of 2, of the received data. In the frequency domain after TEQ and FEQ (*just before slicing*), the model on which these methods rely is the following.

$$\left\{ \begin{array}{l} X^{(0)}[q,t] = H^{(0)}[q]E[q,t] + V^{(0)}[q,t] \\ X^{(1)}[q,t] = H^{(1)}[q]E[q,t] + V^{(1)}[q,t] \end{array} \right\}, \quad q \text{ is the bin rank, and } t \text{ is the time symbol index}$$

Equation 20

15 where:

- $E[q,t]$ is the emitted complex constellation point for the bin q and the symbol t.

- The superscript $^{(i)}, 0 \le i \le 1$, refers to the path.

- $V^{(i)}[q,t]$ stands for the whole perturbation of bin q of symbol t on path i, i.e. the Xtalk and the AWGN.

20 - $H^{(i)}[q]$ is a residual complex "main channel" gain per bin q, on path i.

Equation (19) may be written in a more compact form as:

$X[q,t] = \mathbf{H}[q]E[q,t] + \mathbf{V}[q,t]$, q is the bin rank and t is the symbol time index.

<div align="right">Equation 21</div>

where:

- $\mathbf{R}[q]$ *denotes the correlation matrix of* $\mathbf{V}[q,t]$:
  $$\mathbf{R}[q] = E\{\mathbf{V}[q,t]\mathbf{V}^+[q,t]\}$$ <div align="right">Equation 22</div>

In what follows, we will refer to the **global perturbation model.**

In case of a single Xtalk, the above model reduces to:

$$\begin{cases} X^{(0)}[q,t] = H^{(0)}[q]E[q,t] + G^{(0)}[q]P[q,t] + W^{(0)}[q,t] \\ X^{(1)}[q,t] = H^{(1)}[q]E[q,t] + G^{(1)}[q]P[q,t] + W^{(1)}[q,t] \end{cases} \quad \text{Equation 23}$$

where:

- $W^{(i)}[q,t]$ stands for the AWGN of bin q of symbol t on path i.

- $G^{(i)}[q]$ is a residual complex "coupling channel" gain per bin q, on path i.

- $P[q,t]$ is the Fourier transform of the single Xtalk, per bin q, symbol t, on path i.

Equation (20) may take a more compact form:

$\mathbf{X}[q,t] = \mathbf{A}[q]\mathbf{S}[q,t] + \mathbf{W}[q,t]$, q is the bin rank and t is the symbol time index.

<div align="right">Equation 24</div>

In what follows, we will refer to Equation (23) as the single Xtalk model, and make the

following observations.

- **D1.** Through the symbols, and per bin q the Xtalk $P[q,t]$ is white:

$$E\{P[q,t]P^*[q,t-\tau]\} \propto \delta[\tau] \qquad \text{Equation 25}$$

τ refers to the symbol time index, and not the sampling rate.

26

- **D2.** Under stationarity assumption, and because of the Fourier transform properties, there is no correlation between two different bins of the Xtalk:

$$E\{P[q,t]P^*[q',t]\} \propto \delta[q-q']$$

Equation 26

- **D3.** For all the Xtalks that are finite alphabet in the time domain (HDSL, ISDN, T1), because of the limit central effect of the Fourier Transform, the Xtalk $P[q,t]$ is gaussian, complex and circular.

- **D4.** From one symbol to another one, Xtalk exhibits no correlation (model (1) & (2)):

$$E\{V[q,t]V^+[q,t-\tau]\} = R[q]\delta[\tau]$$

Equation 27

- **D5.** In the case of a single Xtalk and when there is no white noise, the matrix $R[q]$ is singular. In practice however, there is always white noise, $R[q]$ is thus considered as invertible.

Methods considered fall into 2 categories: methods based on the global perturbation model, and methods based on single perturbation model.

Methods based on global perturbation model

Soft maximum Likelihood and slicing : ML estimation and slicing MLES

The method 1) minimizes a weighted norm with respect to the symbol but without any constraint and 2) to slice the relevant result. We have in a compact form:

$$\hat{E}_{MLES}[q,t] = \left[\frac{H^+[q]R^{-1}[q]X[q,t]}{H^+[q]R^{-1}[q]H[q]} \right]_{Slice}$$

Equation 28

This is also known as the covariance matrix. As will be appreciated, $\mathbf{H}^+$ represents the transpose conjugate of $\mathbf{H}$. As in the previous sections, the vector $\mathbf{H}[q]$ and the matrix $\mathbf{R}[q]$ may be estimated blindly, semi-blindly, or in a supervised mode. The blind step involves a decision directed algorithm.

5

Simplified Maximum Likelihood estimation and slicing SMLES

Three other methods (with lower complexity and lower performance) might be deduced from Equation (27):

10 *SMLES1.*

$$\hat{E}_{SMLES1}[q,t] = \left[\frac{\mathbf{H}^+[q]\mathbf{X}[q,t]}{\mathbf{H}^+[q]\mathbf{H}[q]} \right]_{Slice} \qquad \text{Equation 29}$$

SMLES2.

$$\hat{E}_{SMLES2}[q,t] = \left[\frac{\mathbf{1}^+\mathbf{X}[q,t]}{2} \right]_{Slice} \qquad \text{Equation 30}$$

15 where:

$$\mathbf{1}^+ = [1\ 1]$$

This means that the SMLES2 is nothing but an average of the "2 paths".
20

SMLES3 (Brute force method).

$$\hat{E}_{SMLES3}[q,t] = \left[[1\ 0]\mathbf{X}[q,t] \right]_{Slice} \qquad \text{Equation 31}$$

25 As in the previous sections, the vector $\mathbf{H}[q]$ may be estimated blindly, semi-blindly, or in a supervised mode. The blind step involves a decision directed algorithm.

28

Method based on the single Xtalk model g, MINV (matrix inversion method)

The method here consists in applying the invert of the matrix $A[q]$ to $X[q,t]$ and

then, slicing the proper component:

$$\hat{S}[q,t] = A^{-1}[q]X[q,t] \qquad \text{Equation 32}$$

Here the proper component is the first one. We thus have in a compact form:

$$\hat{E}_{MINV}[q,t] = \left[Fist\ component\{A^{-1}[q]X[q,t]\} \right]_{slice} \qquad \text{Equation 33}$$

The matrix $A[q]$ can be estimated in a supervised, semi-blind or full blind mode, as

usual. The blind step involves a decision directed approach.

The foregoing description has been presented for purposes of illustration and

description. It is not intended to be exhaustive or to limit the invention to the precise

forms disclosed. Obvious modifications or variations are possible in light of the above

teachings. The embodiment or embodiments discussed were chosen and described to

provide the best illustration of the principles of the invention and its practical application

to thereby enable one of ordinary skill in the art to utilize the invention in various

embodiments and with various modifications as are suited to the particular use

contemplated. All such modifications and variations are within the scope of the invention

as determined by the appended claims when interpreted in accordance with the breadth to

which they are fairly and legally entitled.